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Simultaneous wideband transmission of five FDM signals over a fiber optic link

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THESIS

SIMULTANEOUS WIDEBAND TRANSMISSION
OF FIVE FDM SIGNALS OVER A
FIBER OPTIC LINK

by

Ilias K. Dimopoulos

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19 Abstract (continue on reverse if necessary and identify by block number)

The purpose of this project was the design, construction and testing of a fiber optic communication link for the simultaneous transmission of five audio signals using the technique of frequency division multiplexing. The five subcarrier frequencies were between 1 and 10 MHz. Amplitude modulation with two sidebands and suppressed carrier using an analog voltage multiplier (AD834) was chosen for efficient transmission through the fiber. The unwanted image signals and intermodulation products from the mixing operation were removed using active bandpass filters. An interesting result was the successful design and operation of active filters at the high frequencies using the Generalized Immitance Converter (GIC) configuration. An LED and pin-photodiode pair was used for the optical portion of the link. Simulated link distances of up to 3.2 kilometers were attained using these components. A constant gain for signal frequencies ranging between 200 Hz and 20 kHz was observed during the operation of the system with all the five channels working simultaneously.

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Simultaneous Wideband Transmission of Five FDM Signals Over a Fiber Optic Link

by

Ilias K. Dimopoulos
Lieutenant, Hellenic Navy
B.S., Hellenic Naval Academy, 1980

Submitted in partial fulfillment of the requirements for the degree of

MASTER OF SCIENCE IN ELECTRICAL ENGINEERING

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December 1989

ABSTRACT

The purpose of this project was the design, construction and testing of a fiber optic communication link for the simultaneous transmission of five audio signals using the technique of frequency division multiplexing. The five subcarrier frequencies were between 1 and 10 MHz. Amplitude modulation with two sidebands and suppressed carrier using an analog voltage multiplier (AD834) was chosen for efficient transmission through the fiber. The unwanted image signals and intermodulation products from the mixing operation were removed using active bandpass filters. An interesting result was the successful design and operation of active filters at the high frequencies using the Generalized Immitance Converter (GIC) configuration. An LED and pin-photodiode pair was used for the optical portion of the link. Simulated link distances of up to 3.2 kilometers were attained using these components. A constant gain for signal frequencies ranging between 200 Hz and 20 kHz was observed during the operation of the system with all the five channels working simultaneously.

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ACKNOWLEDGMENTS

I would like to take this opportunity to express my sincere appreciation to the many people who have assisted me in the completion of my thesis and my studies at the Naval Postgraduate School. Specifically, I would like to thank Professor John P. Powers who as my thesis advisor provided me with the knowledge and guidance at critical points of my work. Secondly, I wish to acknowledge John Glenn whose timely support in both technical knowledge and experience as the manager of the fiber optics laboratory, contributed greatly to the success of my thesis. Finally, I would like to express my deep appreciation to my wife Vaso for her support and understanding throughout my endeavors in these two and a half years of studies at NPS.

I. INTRODUCTION

A great deal of work has been done since 1968 in the area of optical communications systems using fiber as the transmission medium. There are many advantages using such a communication link, as seen in [Refs. 1, 2], with the most important being the increased bandwith capability, small size and weight, flexibility, reliability and low cost. The goal of this thesis was to design, build and test a complete fiber communication link for the simultaneous transmission of five audio signals using the principle of Frequency Division Multiplexing (FDM).

Multiplexing is a technique whereby a number of signals can be combined into a composite signal suitable for transmission over a common channel. If this is accomplished by separating the signals in frequency, then the technique is referred to as *Frequency Division Multiplexing* (FDM). In Figure 1 the basic concept of FDM is illustrated using only two channels. Two signals, each covering the frequency band between 0 and 20 kHz, are modulated by a different carrier frequency, namely 2.2 MHz and 3.8 MHz. After the mixer the spectrum of each signal is translated to a subcarrier frequency and, because AM-DSB/SC modulation is used, the bandwidth is twice the baseband. When these two modulated signals are added, we have a composite signal which covers the frequency range from 2.18 to 3.82 MHz, utilized a total of 1.64 MHz.

In this work only purely analog techniques were used. The main reasons for this choice of analog techniques are:

- Less complexity of the circuitry due to the lack of a digitization process, hence no need for analog-to-digital converters.
- Lower cost due to the lack of coding equipment.
- Avoidance of coding errors and therefore coding noise in the circuit.

Another reason to explore the capabilities of analog transmission through the fiber comes for the possible applications of such a system, which are:

- Analog fiber transmission has a part to play in communication networks, especially in situations where the optical fiber link is part of a larger analog network.

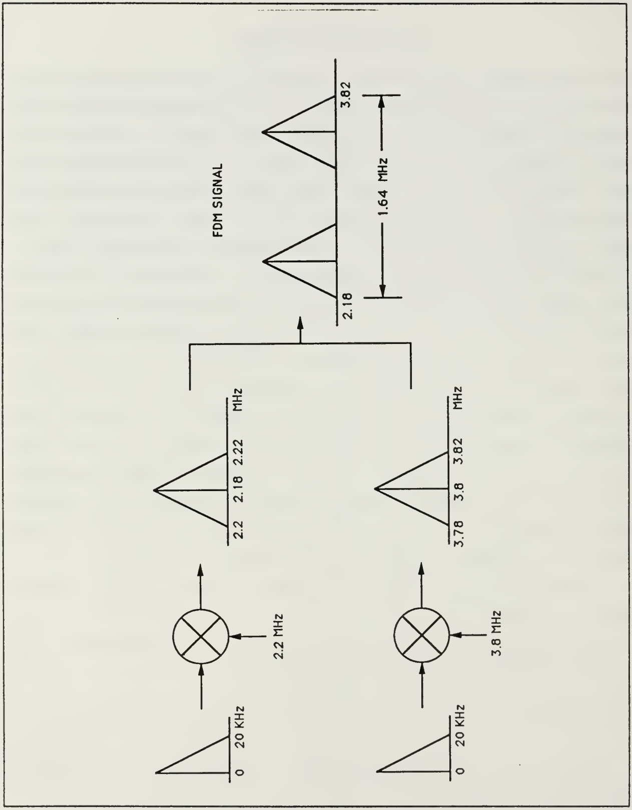

Figure 1. The Concept of Frequency Division Multiplexing.

- There are applications such as direct cable television and common antenna television (CATV) where analog systems may be utilized.
- An analog system can be easily converted to digital by using an A/D converter and a digital signal processor.
- Analog techniques are presently used for underwater data transmission and sensors.

To separate the five channels in frequency, bandpass filters had to be used. Trying to find an active filter configuration that could work at frequencies above 1 MHz, the author was very pleased to find that, using a video amplifier and the Generalized Immitance Converter (GIC), he could design active filters that worked up to 10 MHz. This frequency range had previously not been achieved with this design.

While each of the components of this system will be discussed in detail, Figure 2 shows a block diagram of a single channel. A message with 20 kHz bandwidth is used to modulate a subcarrier with a fixed frequency between 2 and 10 MHz. Double sideband/suppress carrier (AM-DSB/SC) modulation was chosen. The component used to produce this modulation was an analog voltage multiplier (AVM) or mixer. Following the mixer, a bandpass filter was used to remove any signal artifacts that were outside the channel. A summer combined the modulated subcarriers from the other channels. An electrical-to-optical converter launched the light into the fiber and an optical-to-electrical converter produced a signal for the demodulation process. A bandpass filter was used again to separate the subcarrier of interest out of the FDM signal. Another mixer provided coherent demodulation, and finally, a low pass filter recovered the original information.

A block diagram of the whole system with the three main parts (the transmitter, the channel (fiber) and the receiver) can be seen in Figure 3. Five different mixers were used to produce the AM-DSB/SC signals followed by five bandpass filters to remove the signal artifacts outside of the channel. An amplifier in a summing configuration was next for the multiplexing. A DC bias also added and the combined signal was used to drive a HFBR-1404 LED and to intensity modulate the light for the efficient transmission through the fiber.

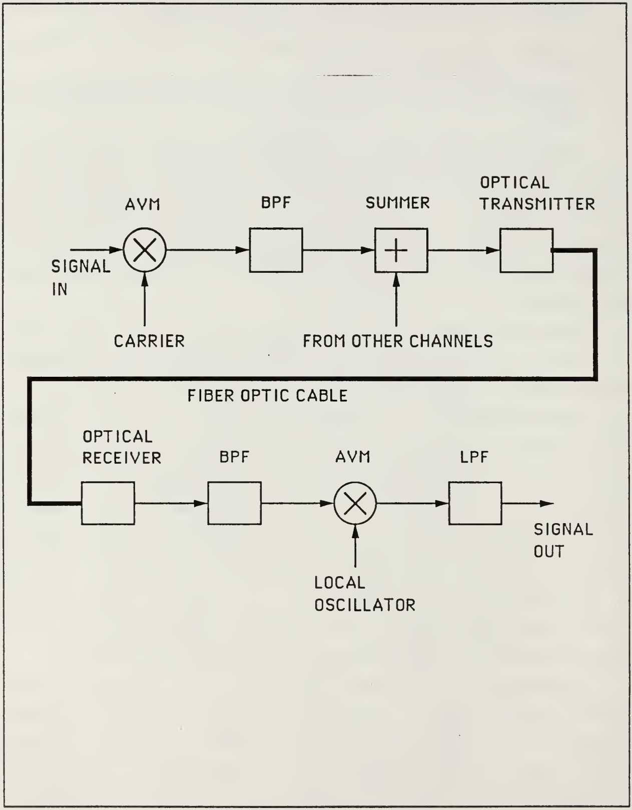

Figure 2. The Communication Link with Only One Channel.

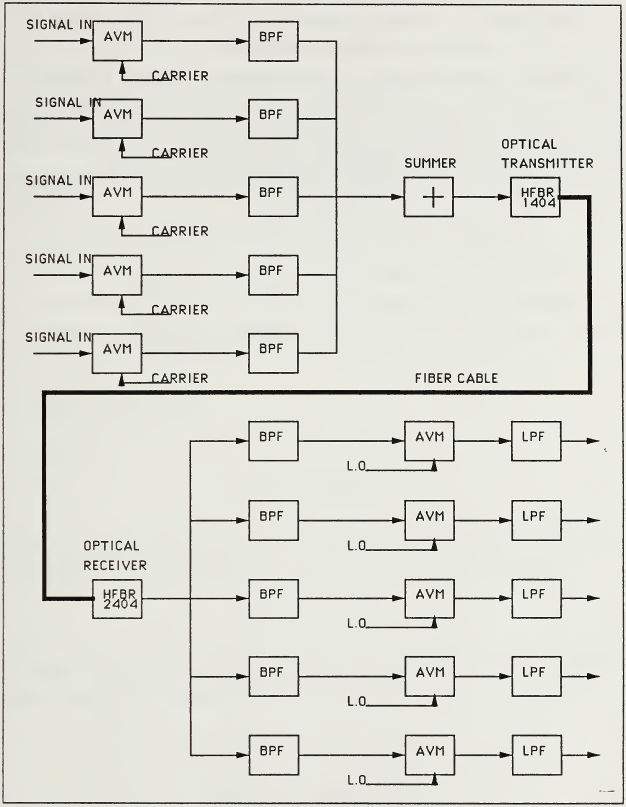

Figure 3. Block Diagram of the Communication Link.

At the other end of the fiber cable, the HFBR-2404 optical receiver performs the inverse function of the optical transmitter to reproduce the FDM signal which goes into five parallel bandpass filters for the isolation of the five different channels. The demodulation process was accomplished coherently or synchronously, by first multiplying the AM-DSB/SC wave with a local oscillator signal exactly synchronized in both frequency and phase, and then low-pass filtering the product to recover the original information with the 20 kHz bandwidth.

Having the structure described in Figure 3 in mind, we can now proceed to discuss in detail the different parts of the system.

II. TRANSMITTER

The main purpose of the transmitter is to convert the signal into a form suitable for transmission over the channel. This is achieved, in our case, by amplitude modulation. The block diagram of the transmitter used to modulate each of the five channels and then sum them in frequency is seen in Figure 4 and consists of

- five analog voltage multipliers to produce the AM-DSB/SC signals,
- five active bandpass filters to separate the frequency bands of interest,
- a summing amplifier to accomplish the frequency division multiplexing, and
- the HFBR-1404 optical transmitter to produce light which is intensity modulated by the FDM signal.

A. AM MODULATOR

Consider a sinusoidal carrier wave

$$c(t) = A_c \cos 2\pi f_c t \tag{1.1}$$

where A_c is the carrier amplitude and f_c is the carrier frequency with zero phase. Let m(t) be the message signal, in this case a sinusoidal modulating signal,

$$m(t) = A_m \cos 2\pi f_m t \tag{1.2}$$

with amplitude A_m and frequency f_m. Passing these two signals through the multiplier we obtain an AM-DSB/SC signal, denoted by

$$\begin{aligned} s(t) &= A_c A_m \cos 2\pi f_c t \cos 2\pi f_m t \\ &= \frac{1}{2} A_c A_m \cos 2\pi(f_c + f_m)t + \\ &+ \frac{1}{2} A_c A_m \cos 2\pi(f_c - f_m)t. \end{aligned} \tag{1.3}$$

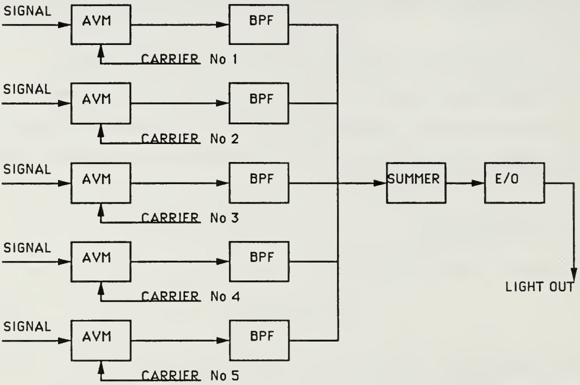

Figure 4. Block Diagram of the Transmitter.

The Fourier transform of s(t) is

$$S(f) = \frac{1}{4} A_c A_m [\delta(f - f_c - f_m) + \delta(f + f_c + f_m) + \delta(f - f_c + f_m) + \delta(f + f_c - f_m)]. \tag{1.4}$$

Therefore, the spectrum of the AM-DSB/ SC wave consists of delta functions at the sum and difference frequencies. Notice also that the transmission bandwidth is twice the bandwidth of the modulating signal m(t). Other cases where the signals are not sinusoids can be found in Refs. [3, 4].

The multiplier selected to produce the AM-DSB/SC wave was the AD834 analog voltage multiplier configured as in Figure 5. There are actually five mixers all connected the same as seen in the previous Figure, with the 20 kHz bandwidth, 300 mV simulated message connected to pin 1 and the carrier connected to pin 8. The center frequencies for the five carriers are

Channel No 1: f_c =2.2 MHz

Channel No 2: f_c =3.86 MHz

Channel No 3: f_c =5.7 MHz

Channel No 4: f_c = 7.3 MHz

Channel No 5: f_c = 9.32 MHz

The AD834 performs very well up to frequencies of 500 MHz but the amplitude of the two input signals must be less than 1 Vp-p. Under this condition, the output is 400 mVp-p maximum. Since the signals in the system had less amplitude than 1 Vp-p, the AM-DSB/SC signal produced by the multiplier needed to be amplified to drive the active bandpass filter. A 1/4 VA4708 operational amplifier configured as a buffer follows the mixer and another 1/4 VA4708 configured as a 10x amplifier increases the signal to drive the filters.

B. TRANSMITTER BANDPASS FILTERS

In order to transmit a number of signals over the same channel, the signals must be kept separated in frequency so that they do not interfere with each other and so that they can be separated at the receiving end. In practice, the actual spectrum of a signal may be much wider than the nominal spectrum which means

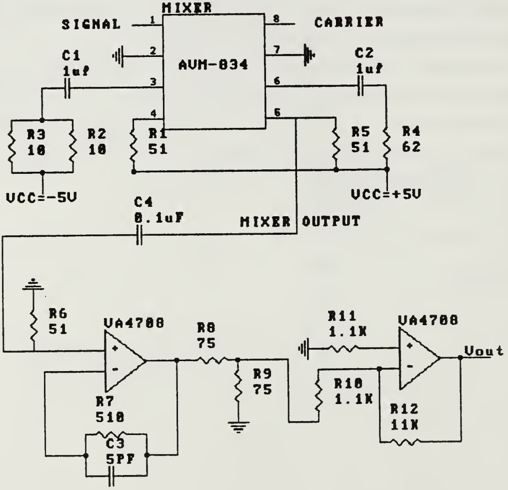

Figure 5. Analog Voltage Multiplier Configuration.

that some interference will occur among adjacent channels. For this reason a bandpass filter must be provided in every channel to ensure separation between the channels.

It was expected that the design of active filters would be a difficult task, since the frequencies of interest were above 1 MHz and it is well-known that the greatest benefits from the use of active components in electrical filter networks are derived for the frequency range f < 500 kHz [Ref. 5: p. 1]. This was really the case since it took more than two months, after trying almost all the designs existed in literature, to get finally the GIC circuit to work as a bandpass filter for frequencies up to 10 MHz.

There are, of course, passive filters which can perform very well up to 500 MHz, so why use active bandpass filter? The advantages are:

- Increased reliability.
- Smaller size which leads to a reduction in parasitics.
- Simpler design and tuning.
- Signal amplification. (In contrast, the passive filter exhibits considerable loss during the operation in the circuit).
- A significant reduction in price.

On the other hand there are some drawbacks of the use of active filter, which are:

- More components are present, therefore there is increased sensitivity to environmental changes.
- Active filter requires power supplies: this means that, in order to conserve energy, it is necessary to design the filter effectively.

Considering the above, it is seen that the advantages outweigh the disadvantages, therefore active filters can be more useful for telecommunications applications. These are the main reasons for the impressive operation of the GIC circuit at high frequencies:

GIC Has a Very Low Active Sensitivity. In practice, real circuit components deviate from their nominal values due to manufacturing tolerances, environmental changes in temperature or humidity and chemical changes. Furthermore,

non-ideal op-amp characteristics and parasitics from capacitor losses can also cause a practical active and passive component to deviate from its ideal behavior. Component deviations of these types cause the circuit transfer function or response to drift away from the specified function. The cause-and-effect relationship between the circuit element variations and the resulting changes in the response or some other network function is referred as *sensitivity*. Ideally, we would like the change in the circuit response to be as small as possible. This implies that either the sensitivity or element variations are small. Small element variations are obtained using precision elements which implies an expensive network. On the other hand, decreasing the sensitivity not only improves the operation of the circuit but the cheaper the circuit becomes to manufacture.

Inductance Simulation. One of the most important applications of the GIC is inductor simulation. This is because low sensitivity active RC filters can be derived directly from passive RLC prototype networks by replacing the passive inductors with active, GIC-simulated inductors. The elimination of inductors characterizes the configuration as an active RC filter and the inductor simulation helps these filters to work at high frequencies.

Amplifier With a High Gain-Bandwidth Product. The invention of video amplifiers with gain-bandwidth product of several hundred MHz has caused a major movement in the successful design and operation of active filters in high frequency applications. The op-amp chosen for this project was the VA4708 quad amplifier provided by VTC Inc. According to the manufacturer's specifications [Ref. 6] the high slew rate (90 V/ μsec), the open loop gain (10 V/mV) and the fast-settling time (0.1 % in 150 ns) allow the amplifier to be used in analog and high-speed processing applications. Another breakthrough with this amplifier is the fact that it operates with nominal supply voltages of +5 and −5 volts. This scaling in supply voltages from the old +15 and −15 volts standard to the lower values reduces the power consumption.

Figure 6 is an illustration of a practical bandpass filter realization using single op-amp GIC. Every bandpass filter is characterized by its center frequency f_c, the 3-dB bandwidth B and the quality factor Q defined as

$$Q = \frac{f_c}{B} \, . \tag{1.5}$$

With the GIC one can easily determine these parameters just by choosing the right values for the resistors and capacitors of the circuit, according to the following expressions:

$$R1 = R2 = R3 = R4 = R \tag{1.6}$$

$$C1 = C2 = C \tag{1.7}$$

$$f_c = \frac{1}{2\pi RC} \tag{1.8}$$

$$Q = \frac{R5}{R} \, . \tag{1.9}$$

Another thing to note before describing the actual filters is that, to decrease Q, we just have to lower the value of R5. But to increase Q, instead of increasing R5 which results in more noise in the circuit, it is better to cascade two, three or more stages of identical GIC's to obtain the desired Q. To find out how many stages must be cascaded, there is a practical formula [Ref. 7: p. 177]

$$Q = \frac{Q_1}{\sqrt{\sqrt[n]{2} - 1}} \tag{1.10}$$

where Q is the value of the desired quality, Q_1 is the quality factor of a single stage and n is the number of stages.

The design of the five bandpass filters was done in accordance with the above equations. Center frequencies picked to start the design were 3 MHz, 4.5 MHz, 6 MHz, 8 MHz and 10 MHz. Using equation 1.8 we can calculate the right value for resistor R and capacitor C to achieve the above center frequencies. Also a quality factor $Q_1 = 10$ for one stage, yields a $Q = 15.53$ for two stages (equation 1.10). Using this value for Q, the 3-dB bandwidth can be calculated solving

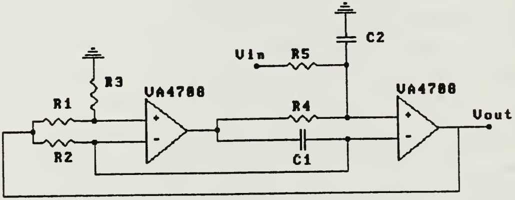

Figure 6. The GIC Configuration.

1.10). Using this value for Q, the 3-dB bandwidth can be calculated solving equation (1.5) for each channel. Figure 7 provides a complete schematic of the 4th-order, 2-stage GIC bandpass filter used in the transmitter and Table 1 gives the calculated values for the passive components for each of the five channels.

In practice the 20% tolerance in the values of resistors and capacitors, as well as the parasitics from the use of "bread-boards" in high frequencies affect the characteristics of the filter, which result in different values for center frequencies and bandwidth than those designed. Figures 8-17 show the measured frequency response gain and phase characteristic plots for each channel with data gathered with the HP 3575A gain-phase meter. A discussion of each channel is presented below.

Channel 1 (Figures 8, 9) has a center frequency at 2.2 MHz and a 3-dB bandwidth of 220 kHz. The output signal is less than the input by 1 dB and the phase characteristic is quite linear with a zero phase shift at the center frequency.

Channel 2 with center frequency 3.86 MHz and a 3-dB bandwidth of 350 kHz (Figures 10, 11) exhibits a gain of 2 dB and a linear phase response in the passband with zero phase shift at 3.7 MHz.

Channel 3 has a center frequency f_c = 5.7 MHz and a 3-dB passband of 400 kHz (Figures 12, 13). The gain at the center frequency is 10 dB and zero phase shift occurs at 6.2 MHz.

Channel 4 has also a high gain of 7.4 dB at the center frequency which is 7.3 MHz. A symmetrical gain plot and a linear phase can be seen in Figures 14 and 15. The bandwidth at 3-dB points is 420 kHz.

Channel 5 has a center frequency f_c = 9.32 MHz, a 3-dB bandwidth B=400 kHz, a gain of 8.6 dB at center frequency and zero phase shift occuring at the same frequency (Figures 16, 17).

Finally, Figure 18 can give a complete view of the location of each frequency response gain plot in the frequency range of 1-12 MHz.

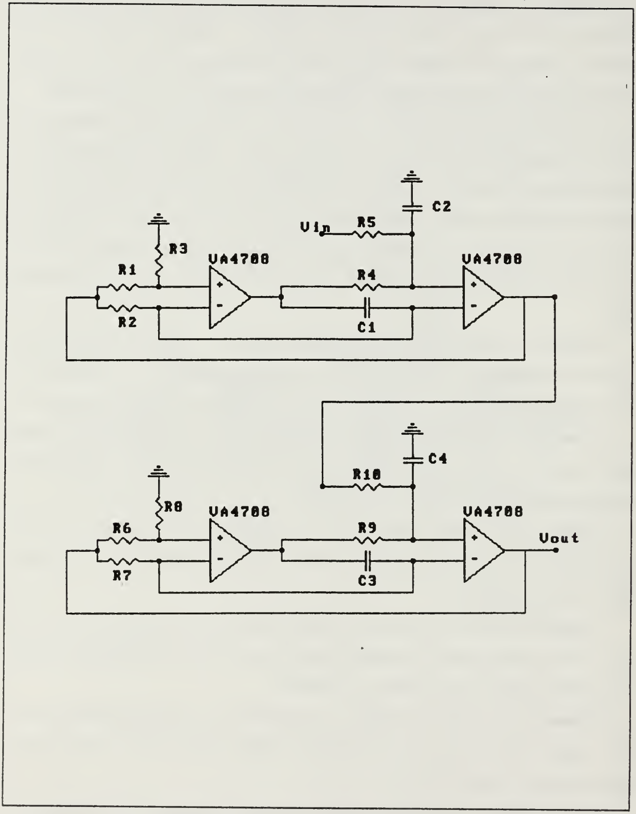

Figure 7. Transmitter Bandpass Filter Configuration.

Table 1. TRANSMITTER BANDPASS FILTER COMPONENT VALUES.

Channel 1	Channel 4
Stage 1	*Stage 1*
R1 = R2 = R3 = R4 = 1.8K	R1 = R2 = R3 = R4 = 510 Ohms
C1 = C2 = 30pF	C1 = C2 = 38pF
Q Resistor R5 = 18K	Q Resistor R5 = 5.1K
Stage 2	*Stage 2*
R6 = R7 = R8 = R9 = 1.8K	R6 = R7 = R8 = R9 = 510 Ohms
C3 = C4 = 30pF	C3 = C4 = 38pF
Q Resistor R10 = 18K	Q Resistor R10 = 5.1K
Channel 2	**Channel 5**
Stage 1	*Stage 1*
R1 = R2 = R3 = R4 = 1.1K	R1 = R2 = R3 = R4 = 510 Ohms
C1 = C2 = 32pF	C1 = C2 = 31pF
Q Resistor R5 = 11K	Q Resistor R5 = 5.1K
Stage 2	*Stage 2*
R6 = R7 = R8 = R9 = 1.1K	R6 = R7 = R8 = R9 = 510 Ohms
C3 = C4 = 32pF	C3 = C4 = 31pF
Q Resistor R10 = 11K	Q Resistor R10 = 5.1K
Channel 3	
Stage 1	
R1 = R2 = R3 = R4 = 820 Ohms	
C1 = C2 = 32pF	
Q Resistor R5 = 8.2K	
Stage 2	
R6 = R7 = R8 = R9 = 820 Ohms	
C3 = C4 = 32pF	
Q Resistor R10 = 8.2K	

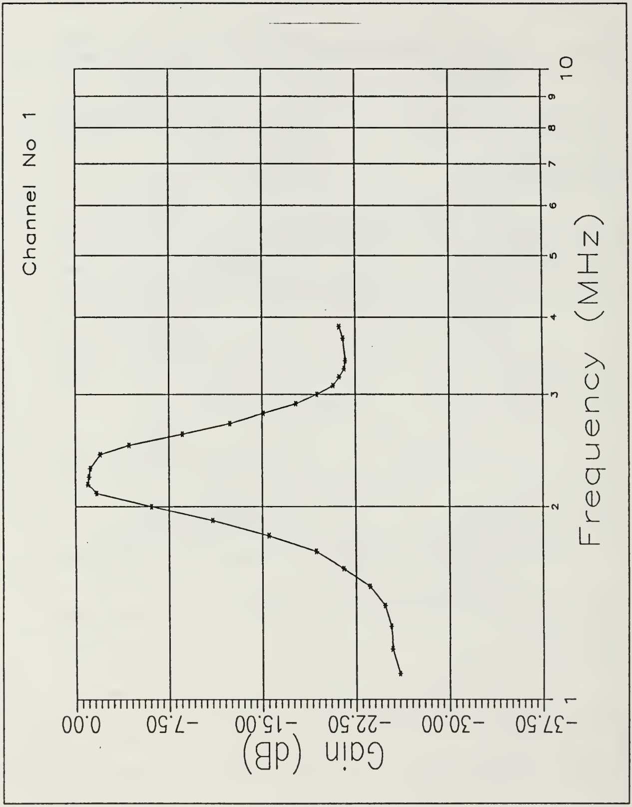

Figure 8. Gain Characteristic of BPF (Channel 1).

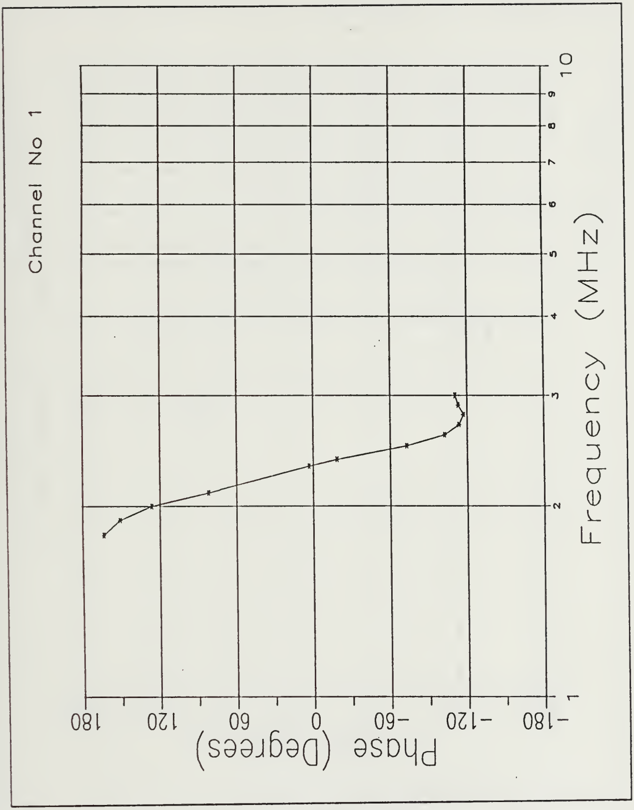

Figure 9. Phase Characteristic of BPF (Channel 1).

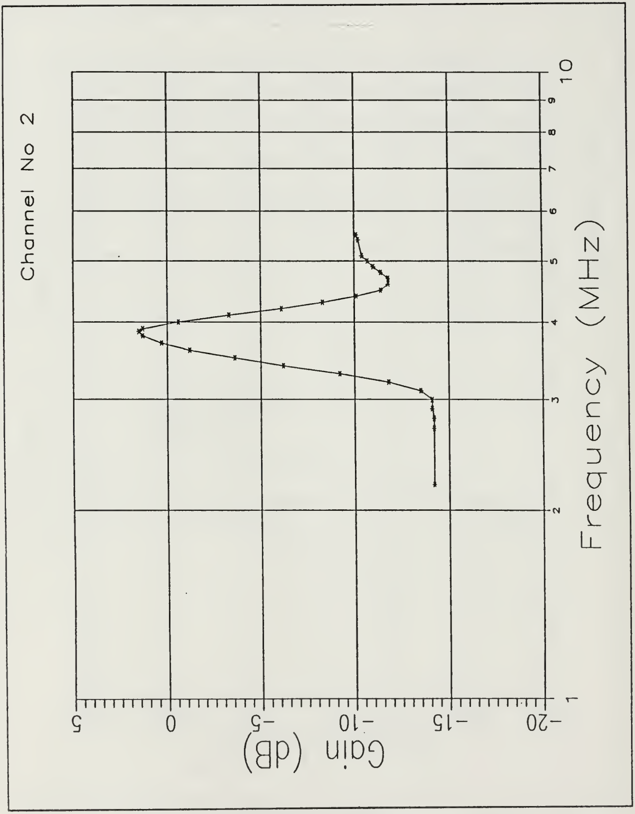

Figure 10. Gain Characteristic of BPF (Channel 2).

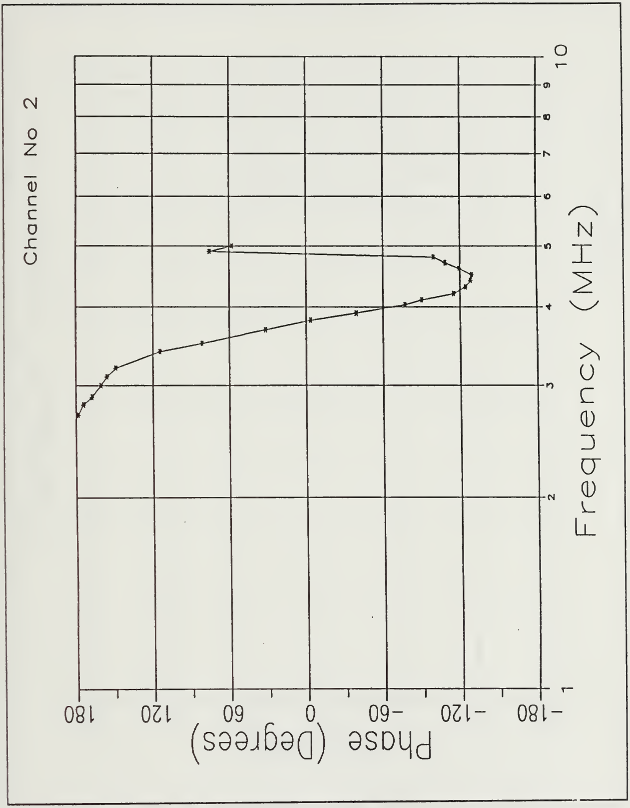

Figure 11. Phase Characteristic of BPF (Channel 2).

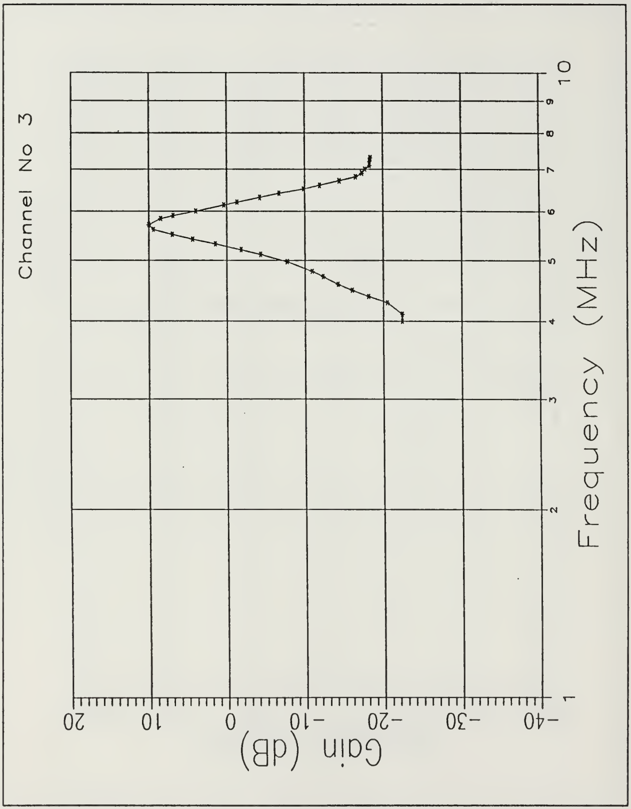

Figure 12. Gain Characteristic of BPF (Channel 3).

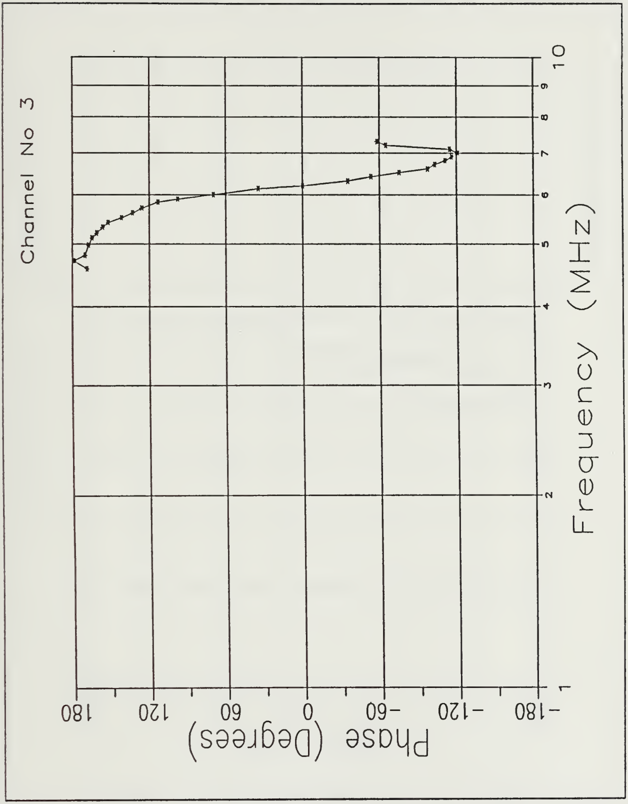

Figure 13. Phase Characteristic of BPF (Channel 3).

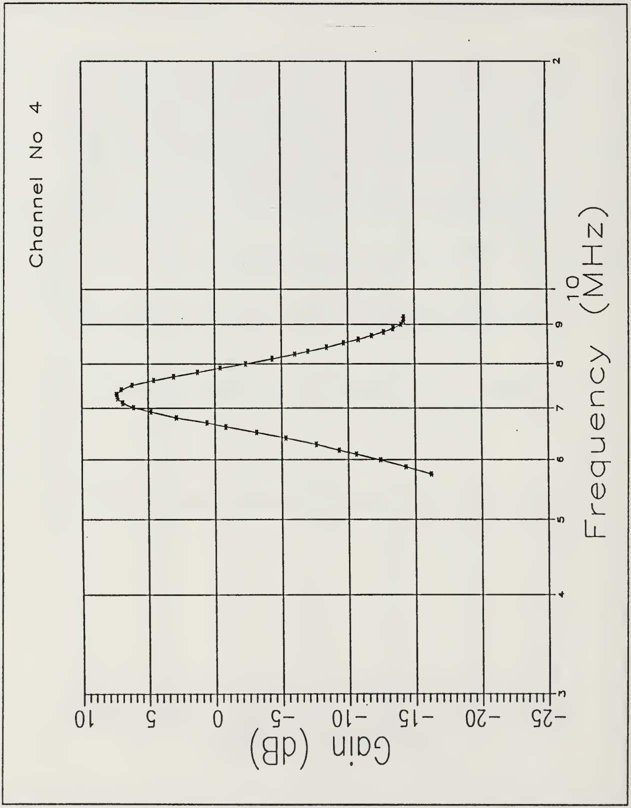

Figure 14. Gain Characteristic of BPF (Channel 4).

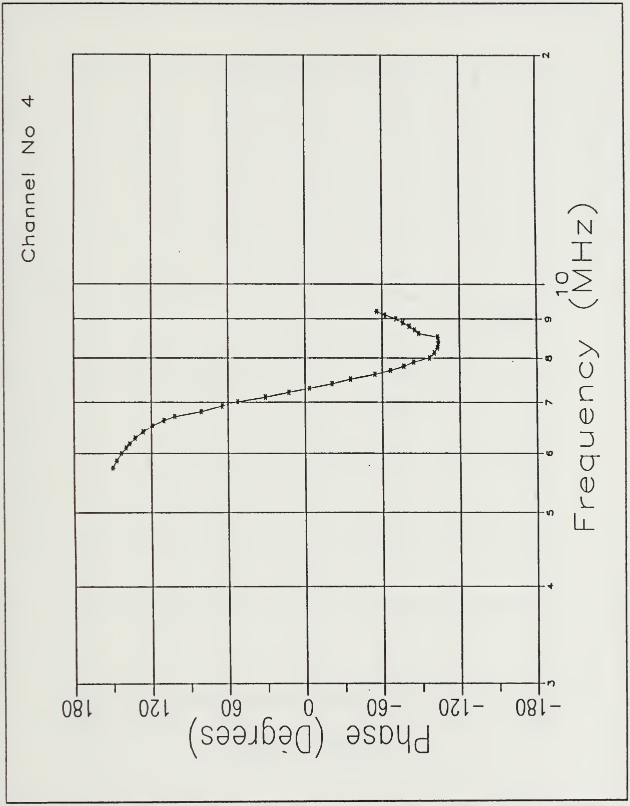

Figure 15. Phase Characteristic of BPF (Channel 4).

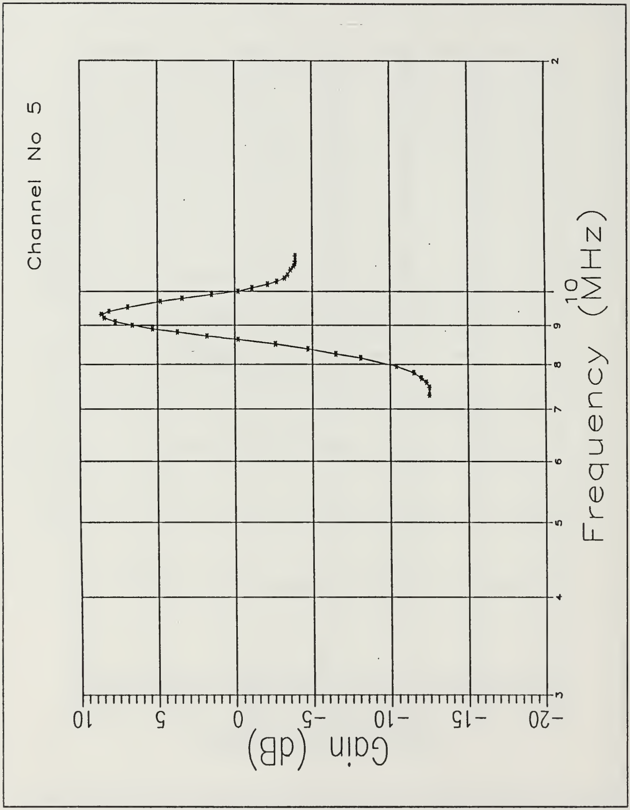

Figure 16. Gain Characteristic of BPF (Channel 5).

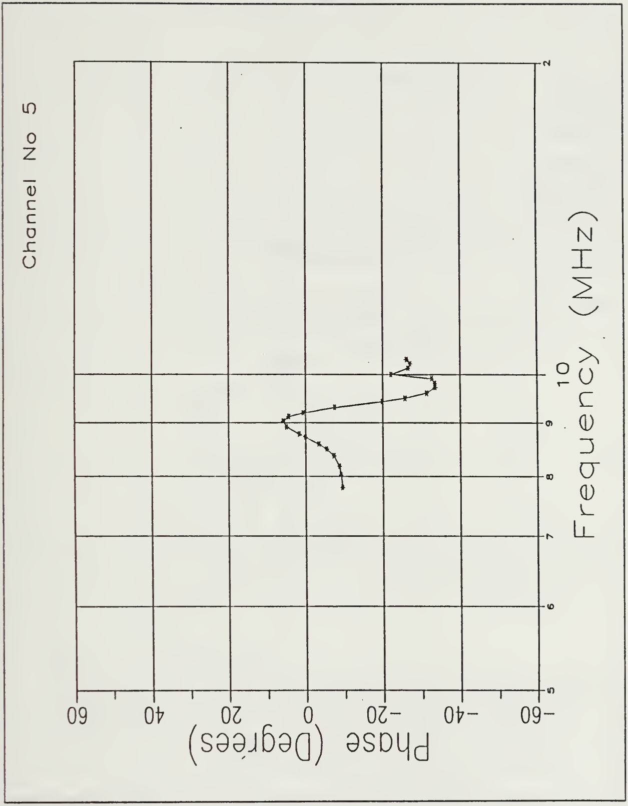

Figure 17. Phase Characteristic of BPF (Channel 5).

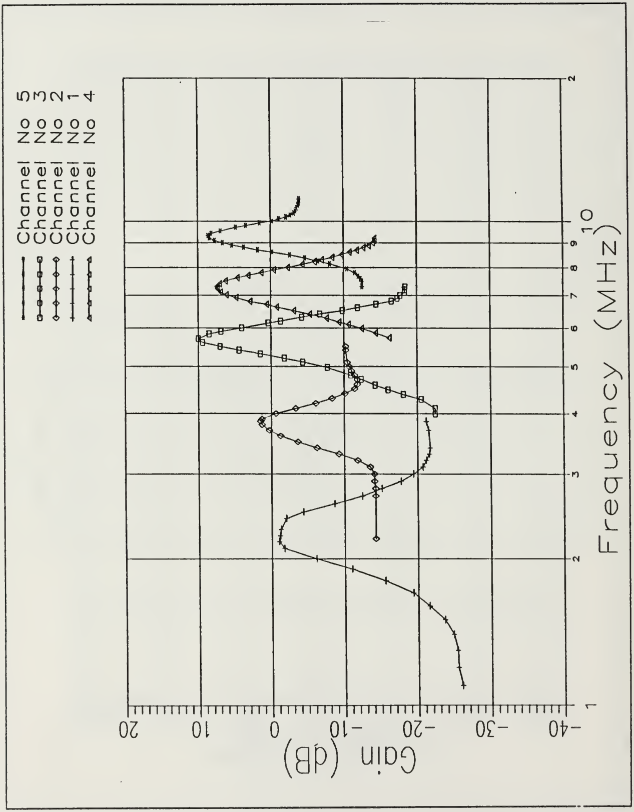

Figure 18. Gain Characteristic of All Five Channels.

C. FREQUENCY DIVISION MULTIPLEXING

Signal multiplexing allows transmission of many signals on the same channel. In frequency division multiplexing (FDM) these signals share the channel but are in separate frequency ranges. Figure 19 illustrates this procedure in the frequency domain, where the signal with 20 kHz bandwidth is modulated by a different subcarrier with values shown in this Figure. At the output of the BPF every modulated signal has a spectrum centered at the frequency of the carrier with a bandwidth that is twice the bandwidth of the original message. For transmission through the fiber, the five signals must be added to give a composite signal covering the frequency band from 2.18 MHz to 9.34 MHz. Thus a total bandwidth of 7.16 MHz must be utilized.

In this thesis, FDM was accomplished using a 1/4 VA4708 amplifier configured as a summer, as shown in Figure 20. The equation which governs the operation of the summer is

$$V_o = -\left\{ \sum_{i=1}^{6} \frac{R_7}{R_i} \times V_i \right\} \qquad (1.11)$$

where V_i, $i = 1,2,...5$ represents the amplitude of the five signals going into the summer and V_6 is the DC bias voltage necessary to bring the amplitude of the signal out of the summer in the range between 1.5 and 2.1 volts in order to drive the optical transmitter.

D. OPTICAL TRANSMITTER

The optical transmitter used in this project to convert the electrical signal to an optical one is the HFBR-1404 transmitter, with a typical output power of −17.5 dBm and a rise/fall time of 4 nsec. The transmitter connections can be seen in Figure 21. In order for the planar 820 nm GaAlAs emitter to be driven, the forward input current must be less than 60 mA. According to the table of the characteristics in [Ref. 8: p. 4], the forward voltage will be between 1.58 and 2.19 volts, with typical value 1.8 volts. This means that a DC bias voltage should be

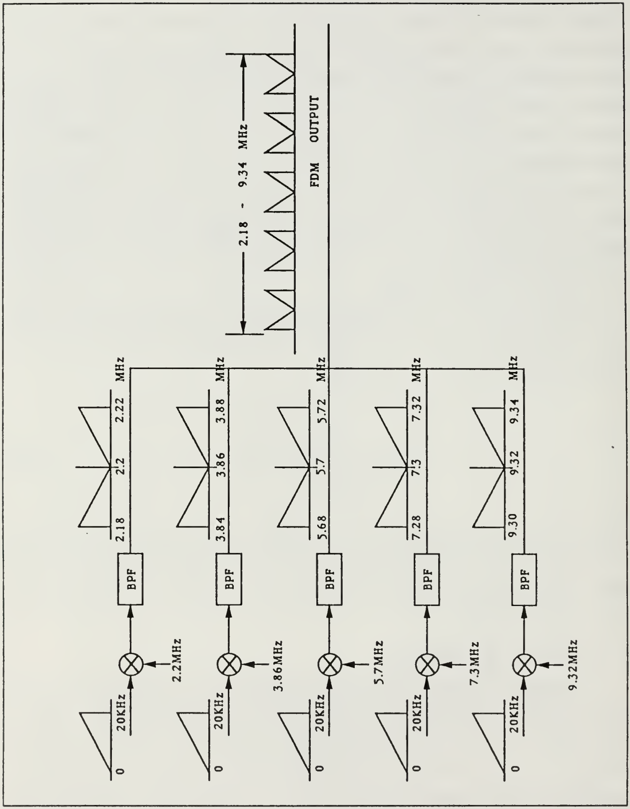

Figure 19. Frequency Division Multiplexing.

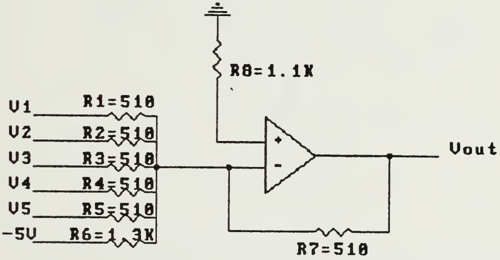

Figure 20. Summing Amplifier Configuration.

voltage should be applied to the summer as mentioned before. The point where the transmitter emits light was found experimentally by changing the value of the resistor R6. Applying the five signals simultaneously, the optical transmitter can be driven into saturation, a condition cured by lowering the amplitude of each signal down to approximately 100 mV. This cannot be done just by lowering the amplitude of the signal out of every function generator because the signal then is not strong enough to drive the transmitter bandpass filters. So, an attenuating amplifier was placed at the output of the BPF in every channel to lower the amplitude of the signal.

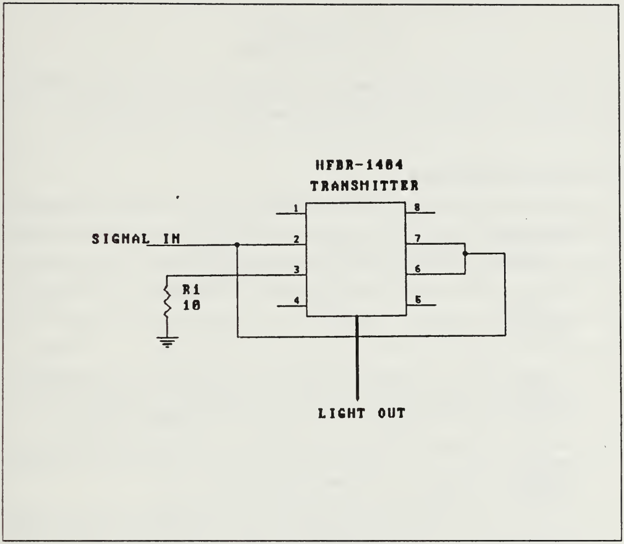

Figure 21. Optical Transmitter Configuration.

III. RECEIVER

Figure 22 shows the block diagram of the receiving part. In every path there is a bandpass filter followed by an AM demodulator and a low-pass filter.

A. OPTICAL RECEIVER

Since the system conveys analog information and the transmitter chosen was the HFBR-1404, the compatible HFBR-2404 fiber optic receiver was used. This contains a discrete PIN photodiode and a preamplifier IC and is compatible with most fiber optic cables. A schematic of the entire configuration of the optical receiver can be seen in Figure 23. Some notes on the operation of this receiver are as follows:

- It is essential that a by-pass capacitor (0.1 μF ceramic) be connected from pin 6 (V_{cc}) to ground. The power supply is connected to this pin in contrast with the similar HFBR-2402 receiver, where it is connected in pin 2.
- After the conversion of the optical signal back to an electrical one, a capacitor is needed in the path for DC isolation, followed by an amplifier connected as a buffer for impedance isolation.
- Since there is a considerable attenuation of the FDM signal passing through the fiber (almost half the transmitting signal), two or three amplification stages (depending on the channel) are needed in order for the signal to drive the filters. However, the two amplification stages shown in the bottom of Figure 23 are common to all the channels. In this way a signal with amplitude 250 mV is obtained out of the optical receiver.

B. RECEIVER BANDPASS FILTERS

Separation of the modulated signal in each channel requires the presence of five parallel bandpass filters centered on the appropriate frequency with a bandwidth corresponding to the channel of interest.

In our case a selection of the GIC configuration identical to that of the transmitter was in order. Since this configuration has a very low active sensitivity, the values of the passive components can be slightly different without any effect on the operation of the filters. A problem arises during the operation because

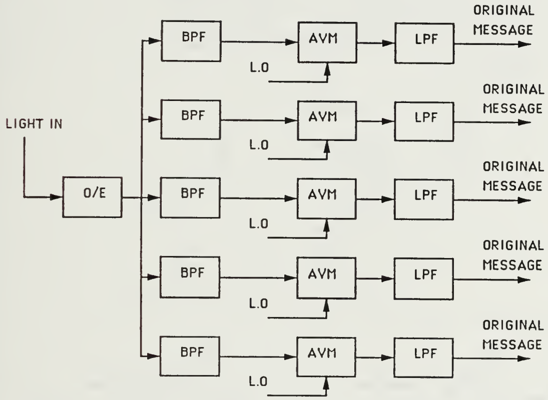

Figure 22. Block Diagram of the Receiver.

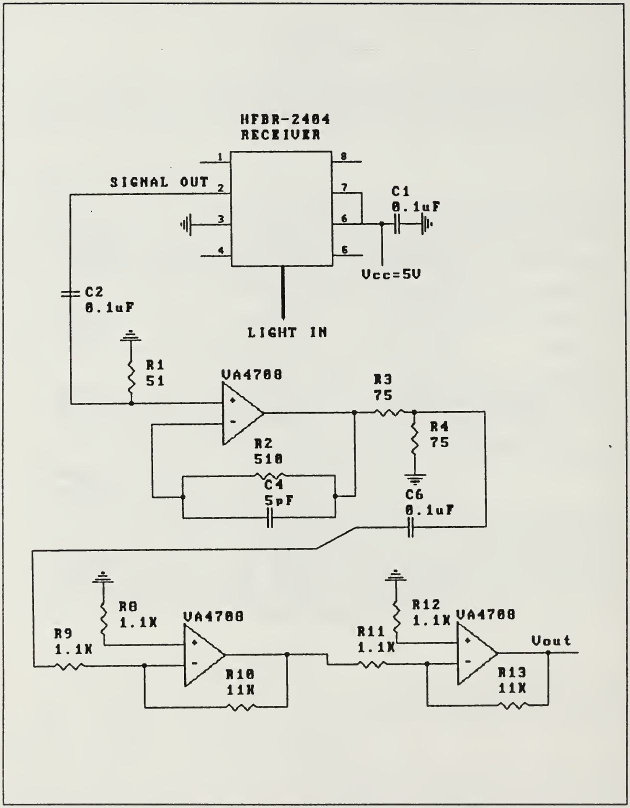

Figure 23. Optical Receiver Configuration.

there is a need to avoid saturation of the filters at the three higher frequency channels. This phenomenon occurs when the amplitude of the signal into the filter is large and above the dynamic range of the filter. Saturation was observed for levels 300 mV or higher. At this point instead of using an attenuating amplifier before the filter, it was found experimentally that the presence of a resistor connected to the non-inverting input of the second amplifier of the filter reduced the gain and still drove the filter properly. Figure 24 illustrates the circuitry of the receiver GIC; the presence of resistors R11 and R12 provides the unsaturated operation of the filter. Table 2 shows the values of the passive components where the appearance of a '0' means that this component is not required for the specific channel.

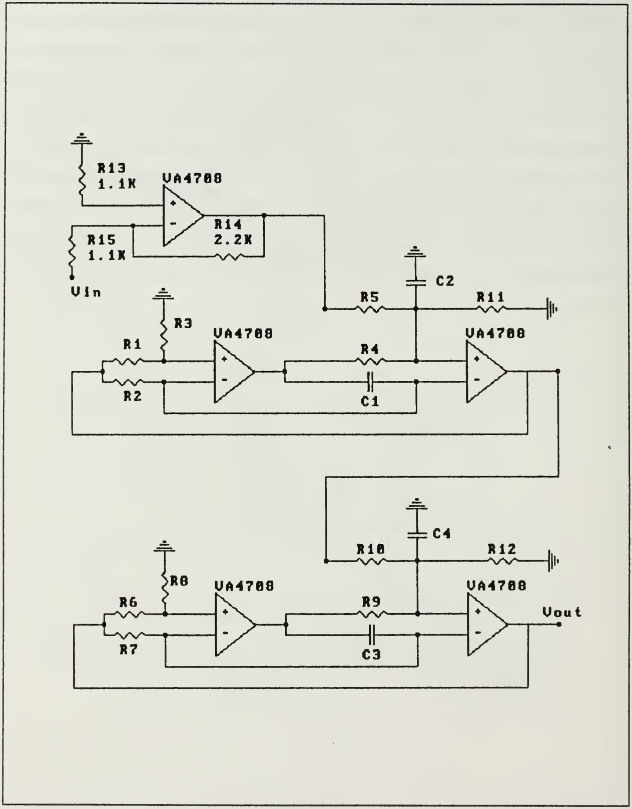

Figure 24. Receiver Bandpass Filter Configuration.

Table 2. RECEIVER BANDPASS FILTER COMPONENT VALUES.

Channel 1	Channel 4
Stage 1	*Stage 1*
R1 = R2 = R3 = R4 = 1.8K, R11 = 0	R1 = R2 = R3 = R4 = 510 , R11 = 11K
C1 = C2 = 30pF	C1 = C2 = 38pF
Q Resistor R5 = 18K	Q Resistor R5 = 5.1K
Stage 2	*Stage 2*
R6 = R7 = R8 = R9 = 1.8K, R12 = 0	R6 = R7 = R8 = R9 = 510 , R12 = 3K
C3 = C4 = 30pF	C3 = C4 = 38pF
Q Resistor R10 = 18K	Q Resistor R10 = 5.1K
Channel 2	**Channel 5**
Stage 1	*Stage 1*
R1 = R2 = R3 = R4 = 1.1K, R11 = 0	R1 = R2 = R3 = R4 = 510, R11 = 1.1K
C1 = C2 = 32pF	C1 = C2 = 31pF
Q Resistor R5 = 11K	Q Resistor R5 = 5.1K
Stage 2	*Stage 2*
R6 = R7 = R8 = R9 = 1.1K. R12 = 0	R6 = R7 = R8 = R9 = 510, R12 = 1.1K
C3 = C4 = 32pF	C3 = C4 = 31pF
Q Resistor R10 = 11K	Q Resistor R10 = 5.1K
Channel 3	
Stage 1	
R1 = R2 = R3 = R4 = 820 , R11 = 11K	
C1 = C2 = 32pF	
Q Resistor R5 = 8.2K	
Stage 2	
R6 = R7 = R8 = R9 = 820 . R12 = 0	
C3 = C4 = 32pF	
Q Resistor R10 = 8.2K	

C. AM DEMODULATOR

The original baseband information can be recovered from the received AM-DSB/SC wave by first multiplying the received signal with a locally generated sine-wave with the same frequency and phase as the subcarrier used in the transmitter and then low-pass filtering the result. The demodulation method is coherent or synchronous detection, since the local oscillator signal is exactly coherent or synchronized in both frequency and phase with the subcarrier wave.

The transmitted signal s(t) is from (1.3)

$$\begin{aligned} s(t) &= \frac{1}{2} A_c A_m \cos 2\pi(f_c + f_m)t \\ &+ \frac{1}{2} A_c A_m \cos 2\pi(f_c - f_m)t. \end{aligned} \tag{2.1}$$

This signal is multiplied with the local oscillator signal

$$c'(t) = A'_c \cos 2\pi f_c \tag{2.2}$$

with amplitude A'_c and frequency f_c, giving

$$\begin{aligned} u(t) &= A'_c \cos 2\pi f_c t \times \frac{1}{2} A_c A_m \cos 2\pi(f_c + f_m)t \\ &+ A'_c \cos 2\pi f_c t \times \frac{1}{2} A_c A_m \cos 2\pi(f_c - f_m)t. \end{aligned} \tag{2.3}$$

Continuing the calculations one step further, we have

$$\begin{aligned} u(t) &= \frac{1}{4} A'_c A_c A_m \cos 2\pi(2f_c + f_m)t \\ &+ \frac{1}{4} A'_c A_c A_m \cos 2\pi f_m t \\ &+ \frac{1}{4} A'_c A_c A_m \cos 2\pi(2f_c - f_m)t \\ &+ \frac{1}{4} A'_c A_c A_m \cos 2\pi f_m t. \end{aligned} \tag{2.4}$$

The first and third terms of frequencies $2f_c + f_m$ and $2f_c - f_m$, respectively, are removed by the low-pass filter leaving

$$u_o(t) = \frac{1}{2} A'_c A_c A_m \cos 2\pi f_m t \tag{2.5}$$

which is analogous to the original modulating wave.

The AD834 is used again to multiply the AM-DSB/SC signal with the local oscillator signal and to give the terms shown in equation 2.4. The signal generators used in the transmitter provided the local oscillator signal for proper synchronization. Figure 25 illustrates the circuitry used for channels 1 and 2. The five signals have different amplitude. In particular, the three higher frequency signals are less than that required to drive the mixer. Hence an amplification stage was necessary for both the AM-DSB/SC and local oscillator signal prior to the input to the mixer for these channels. Thus, for the third channel (Figure 26) a (3x) amplification was used for the AM-DSB/SC signal and (20x) for the local oscillator. In Figure 27 which shows channel 4, a (3x) stage was used again for the signal and a (30x) stage for the local oscillator. Finally, for channel 5 which is the weakest signal, two stages (20x each) were used for the signal and one (10x) stage was used for the local oscillator.

By increasing the amplitude of the signals to a maximum of 1 Vp-p, an output of 250 mV average was obtained for all five channels. Observing equation 2.4, one can see that the demodulator output contains many frequencies; the design of the low-pass filter has to eliminate these frequencies to recover the original information.

D. RECEIVER LOW-PASS FILTER

The circuit chosen to accomplish the low-pass filtering is the basic circuit of the second order multiple feedback (MFB) shown in Figure 29. The transfer function that describes this circuit has the general form

$$\frac{V_2}{V_1} = -K \frac{b_o}{s^2 + b_1 s + b_o} \tag{2.6}$$

where K is the gain of the circuit, b_1 and b_o are constants that can be found from tables in [Ref. 7: pp. 65-73] (depending if the filter type is Butterworth or

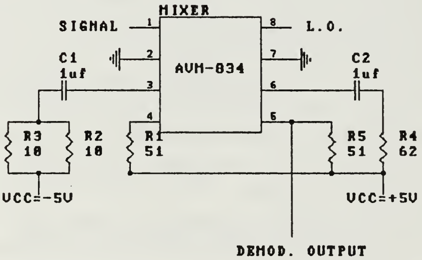

Figure 25. AM Demodulator for Channels 1 and 2.

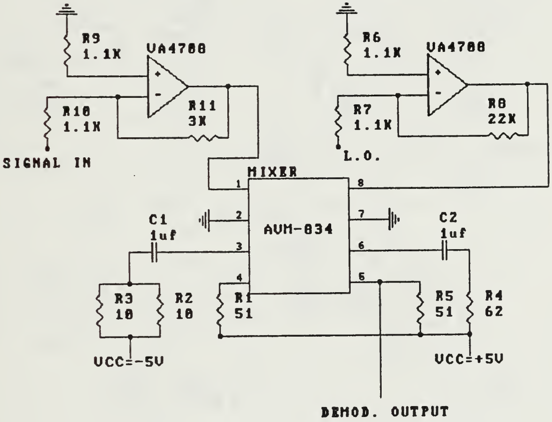

Figure 26. AM Demodulator for Channel 3.

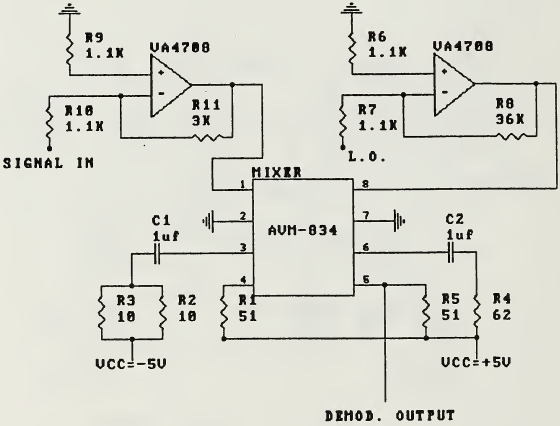

Figure 27. AM Demodulator for Channel 4.

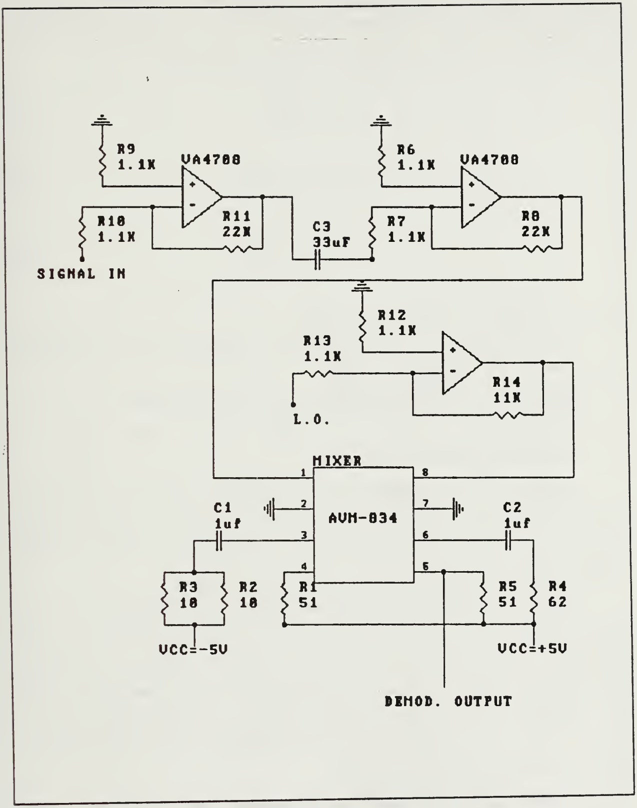

Figure 28. AM Demodulator for Channel 5.

Chebyshev design), V_2 is the filter output voltage, and V_1 is the filter input voltage. The constants of the transfer function K, b_o and b_1 are related to the passive components of the circuit in Figure 29 through the equations

$$b_o = G_2 \frac{G_3}{C_1} \tag{2.7}$$

$$b_1 = G_1 + G_2 + G_3 \tag{2.8}$$

$$K = \frac{R2}{R1} = \frac{G_1}{G_2} \tag{2.9}$$

where $G_1 = \frac{1}{R1}$, $G_2 = \frac{1}{R2}$, and $G_3 = \frac{1}{R3}$.

For the purpose of this design the passive components should be found according to the following requirements:

- The design is for a second order low-pass Butterworth filter.
- The gain is K = 2.
- The cutoff frequency is f_c =50 kHz .

Following the steps outlined in Ref. [7: p .106], with these requirements, a Butterworth low-passfilter was produced with a frequency response shown in Figure 30. As one can see from this Figure, the filter has a very flat response in a large range of frequencies and also a sufficient DC gain.

The actual circuitry for the low-pass filter in every channel can be seen in Figures 31-34. Before proceeding further to the description of the filters, a word of caution should be mentioned about the value of the capacitor C1 which exists after the demodulator and before the amplifier prior to the low-pass filter. Since the message contains frequencies between 200 Hz and 20 kHz, the value of the capacitor at this point is critical, as will be described in a later section about the end-to-end performance of the system.

The amplitude of the signals in every channel is almost the same. In order to study the performance better and to illustrate all five channels on the same plot, a stage with a variable degree of amplification was added prior to the input

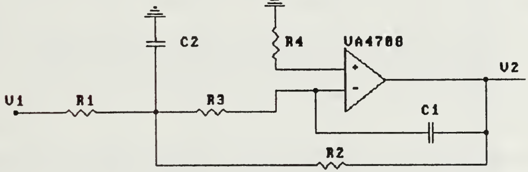

Figure 29. Second Order MFB Low-Pass Filter.

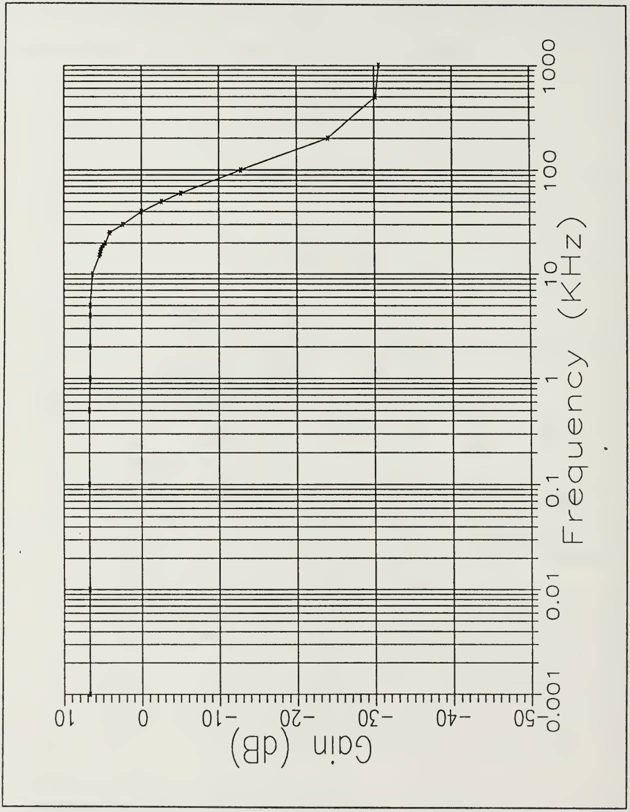

Figure 30. Low-Pass Filter Gain Characteristic.

of each of the low-pass filters. Figure 31 illustrates the configuration for the first channel where a (10x) amplification is used; the signal out of the low-pass filter had an amplitude of 700 mVp-p. Accordingly, two amplification stages (10x and 5x) are seen in Figure 32 for channel 2 giving a signal with an amplitude of 300 mVp-p. Only one stage was needed (10x) for the third and fourth channels (Figure 33) and the output was 150 mVp-p and 100 mVp-p, respectively. Finally, a 180 mVp-p signal was obtained in the fifth channel using a (20x) amplification as illustrated in Figure 34. The frequency of all of the measured signals obtained at the output of the low-pass filter was 20 kHz since this was the frequency used for the experiment, but the system also works for lower frequencies as will be described in section 2 of Chapter 4.

The low-pass filter worked with a cutoff frequency of 25 kHz to pass signals in the entire range from DC to 20 kHz. This is the reason why increasing attenuation was observed as the frequency was increased. Should the frequencies of interest extend beyond 20 kHz, a low-pass filter with greater cutoff frequency must be designed. This can be achieved by multiplying the resistors R1, R2 and R3 by a common factor provided that the capacitors C1 and C2 are divided by the same common factor. This common factor (CF) is calculated from the following equation

$$CF = \frac{f_c}{20\pi} \tag{2.10}$$

where f_c is the new cutoff frequency.

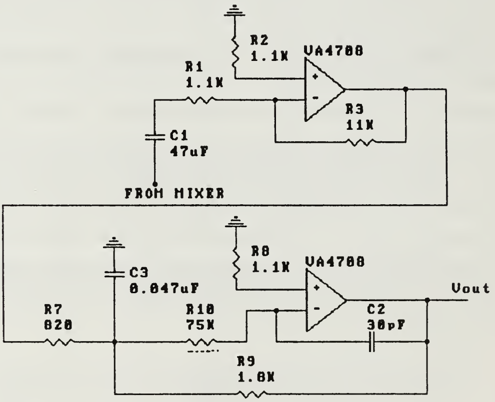

Figure 31. Low-Pass Filter Configuration for Channel 1.

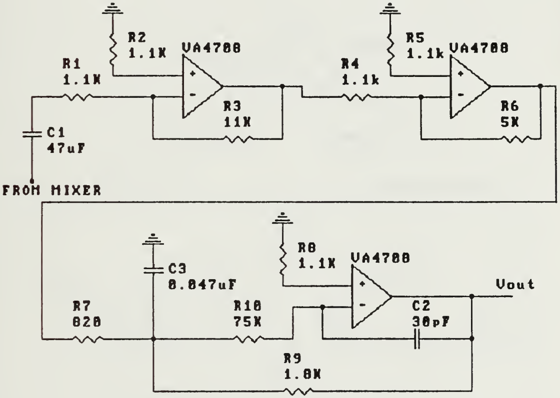

Figure 32. Low-Pass Filter Configuration for Channel 2.

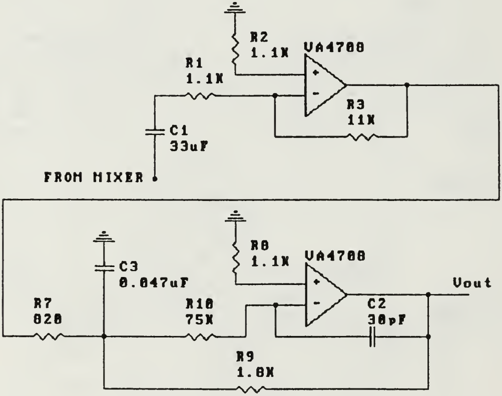

Figure 33. Low-Pass Filter Configuration for Channels 3 and 4.

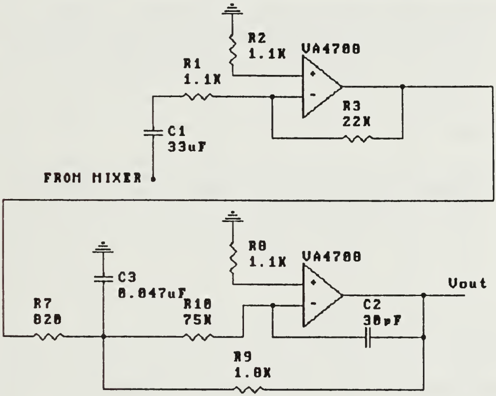

Figure 34. Low-Pass Filter Configuration for Channel 5.

IV. SYSTEM PERFORMANCE

A. ADJACENT CHANNEL INTERFERENCE

Crosstalk or adjacent channel interference occurs when some of the energy of the signal, which is band-limited after the bandpass filter, falls into the adjacent channel, because of the overlapping characteristics of the filters of these channels. One of the reasons for using RF frequencies in this work, was the need for enough separation in frequency between the channels to minimize the effect of crosstalk and for the ability of adding more channels transmitting simultaneously, if so wanted. The active bandpass filters used had a frequency response with a 3-dB bandwidth of 420 kHz average and since the center frequencies of the subcarriers were separated by 2 MHz, more channels could be added between the existing ones. The bandwidth of the filters can be further decreased by cascading two more stages to a value of 90 kHz according equation 1.10. In this way more channels can be added without significant effect from the adjacent channel interference. Although time did not permit to test the system with more than five channels transmitting, this could be a subject for future work.

B. GAIN

The frequencies of interest for transmission ranged from 200 Hz to 4 kHz (voice) and up to 20 kHz. A flat frequency response curve over this range of frequencies for end-to-end transmission would be ideal. Figures 35-39 show that this goal was essentially achieved for all five channels. A different gain for each channel can be seen; this was done for plotting purpose by adjusting the gain of the amplifiers before the low-pass filter in the receiver. Figure 40 illustrates the gain response of all five channels in one plot for comparison.

The value of capacitor C1 at the input of the amplifier in Figures 31 through 34 is critical for the operation of the system in the complete range of frequencies described above. This value is derived from the expression

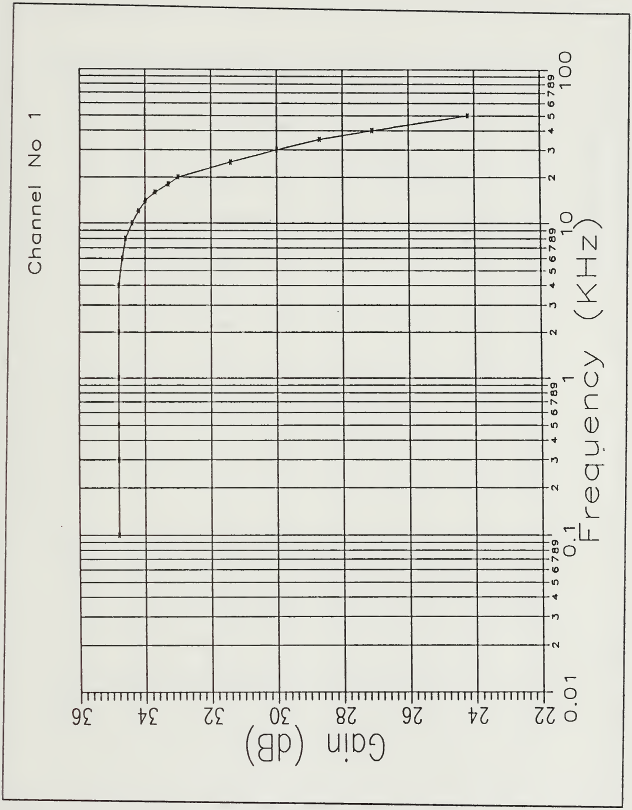

Figure 35. End-To-End Gain Characteristic for Channel 1.

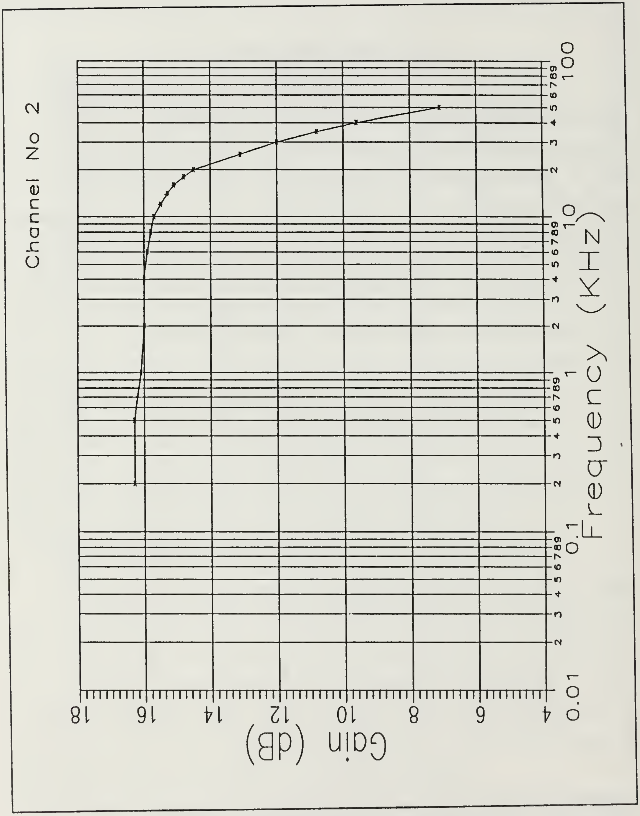

Figure 36. End-To-End Gain Characteristic for Channel 2.

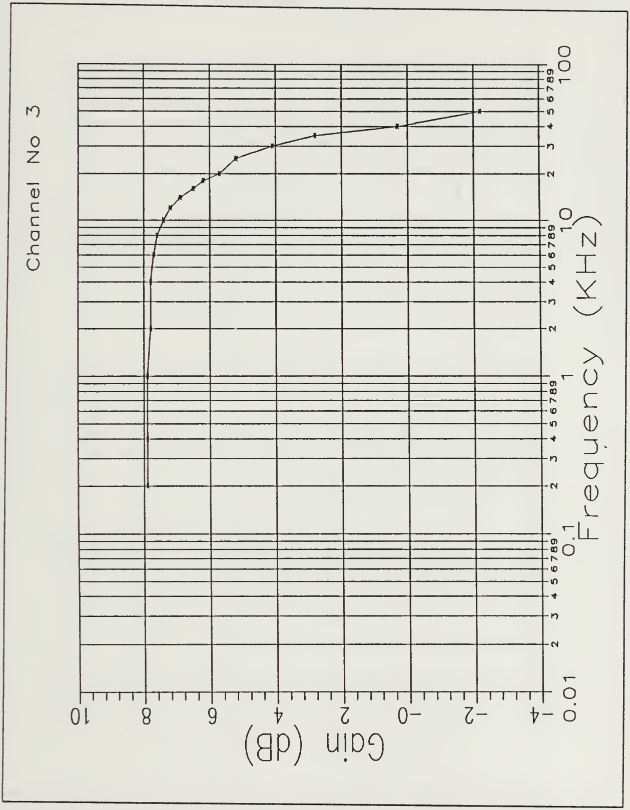

Figure 37. End-To-End Gain Characteristic for Channel 3.

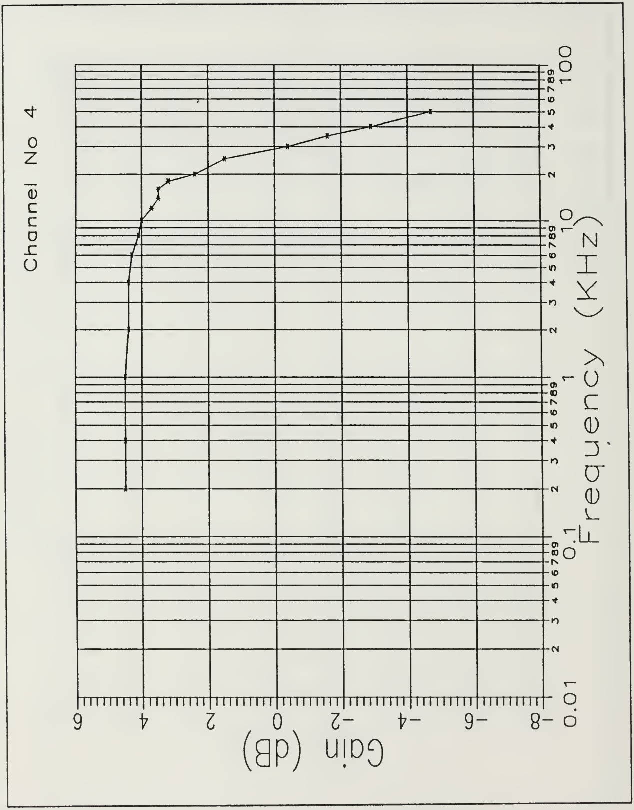

Figure 38. End-To-End Gain Characteristic for Channel 4.

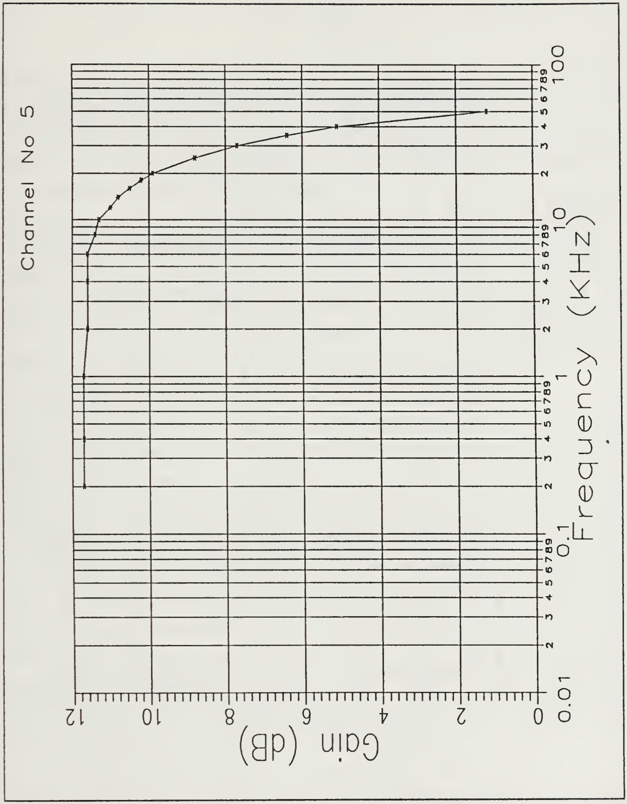

Figure 39. End-To-End Gain Characteristic for Channel 5.

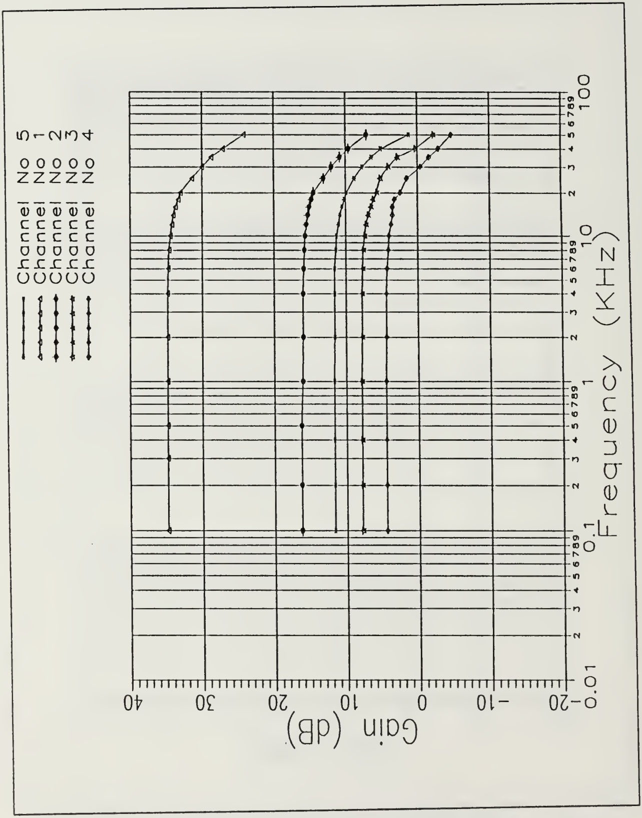

Figure 40. End-To-End Gain Characteristic for All Five Channels.

$$C1 = \frac{1}{2\pi R f} \tag{3.1}$$

where f is the bandwidth of the message. For a signal to be transmitted unattenuated in all the frequencies of interest, the value of the capacitor must be evaluated solving the above equation. For example, choosing a capacitor smaller than 47 μF for channels 1 and 2 (Figures 31 and 32), which seems to be fairly big, causes attenuation at low frequencies.

Although the cutoff frequency of the low-pass filter, by design, should be 50 kHz, the 20% tolerance of resistors and capacitors made the actual filter have a cutoff frequency of 25 kHz. This is satisfactory for the purpose of this design, since the maximum signal frequency transmitted was 20 kHz. The system can easily work with higher frequencies by increasing and decreasing the value of capacitors and resistors of the low-pass filter, respectively, by the same amount.

C. PHASE LINEARITY

Coherent detection was used for the demodulation of the subcarrier at the receiver. This process implies the need of a local oscillator wave synchronized both in frequency and phase with the transmitted modulated wave. If there is a phase error, the demodulated signal $u_o(t)$ contains an unwanted component which cannot be removed by filtering. This component produces distortion appearing as phase distortion. Synchronization was achieved by using the same signal generator for the receiver as in the transmitter. Looking at Figures 41 through 45, which represent the end-to-end phase characteristic of every channel, a fairly linear phase response is observed and thus we can consider a practically distortionless transmission. The phase distortion is usually not serious with voice communications because the human ear is relatively insensitive to phase distortion, but becomes important for data transmission.

D. SYSTEM POWER BUDGET AND RANGE

The minimum required optical power was calculated using a Model 1950 Continuous Optical Attenuator, which is an optomechanical instrument and produces a calibrated optical power loss in dB between the input optical port and

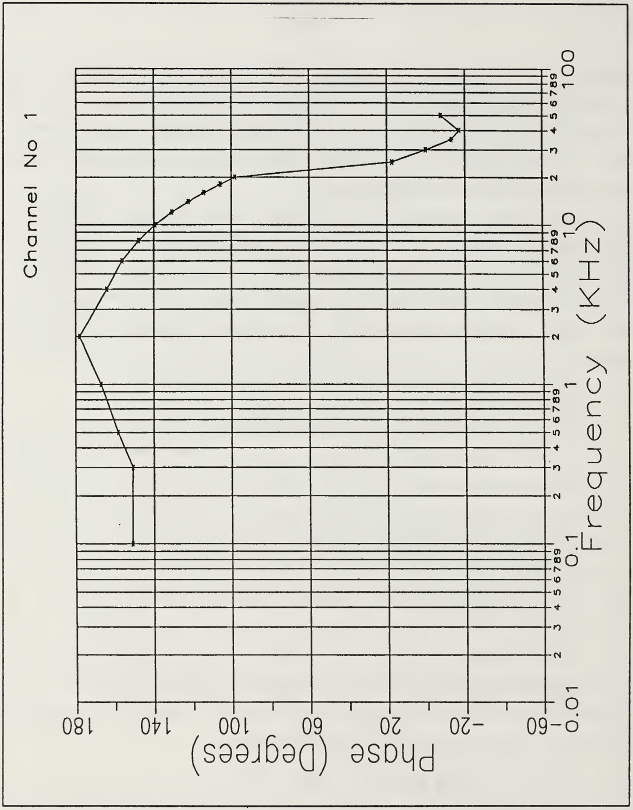

Figure 41. End-To-End Phase Characteristic for Channel 1.

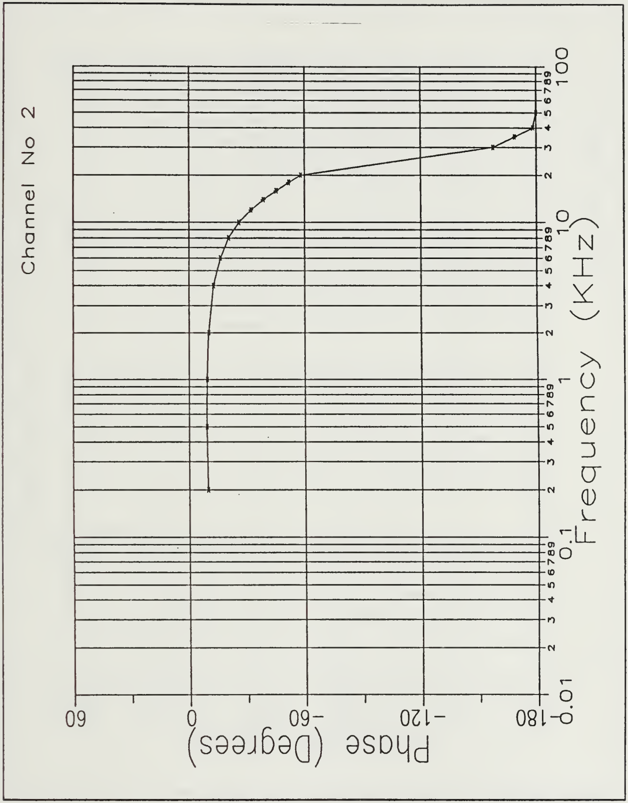

Figure 42. End-To-End Phase Characteristic for Channel 2.

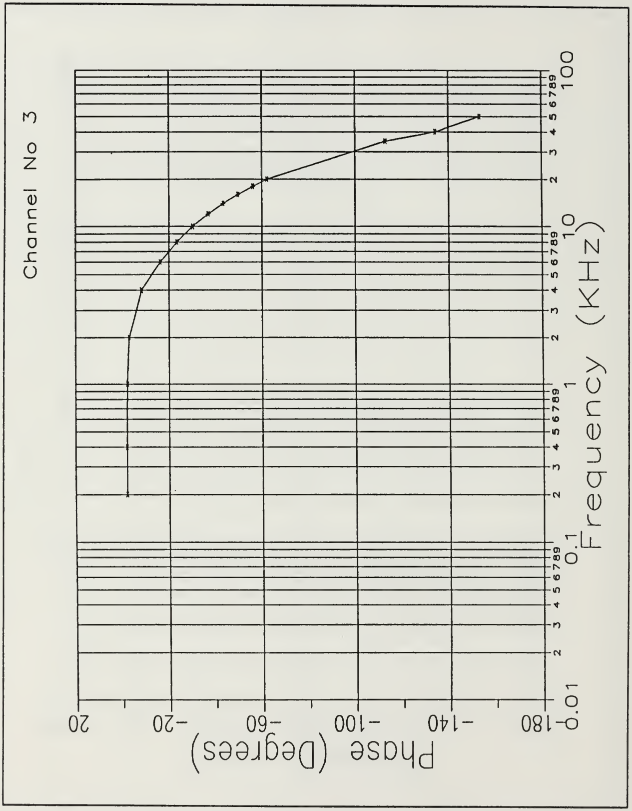

Figure 43. End-To-End Phase Characteristic for Channel 3.

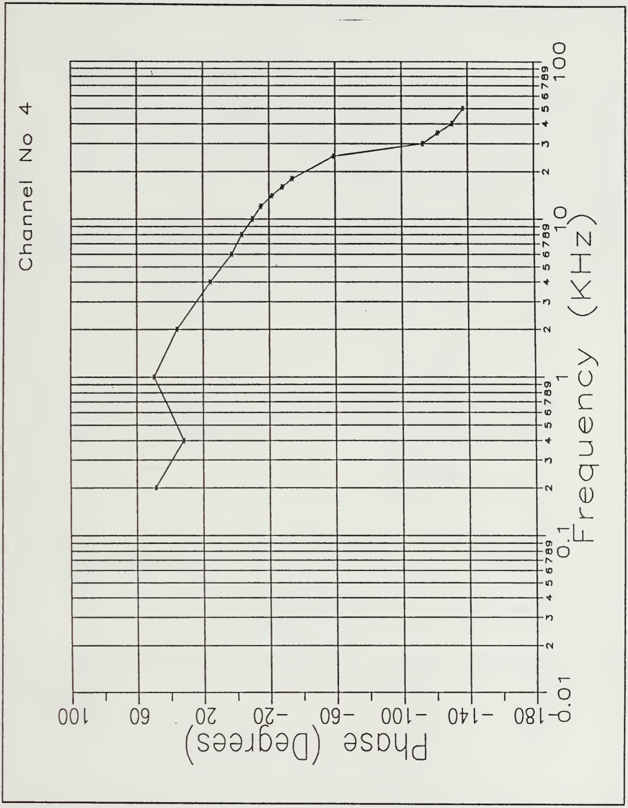

Figure 44. End-To-End Phase Characteristic for Channel 4.

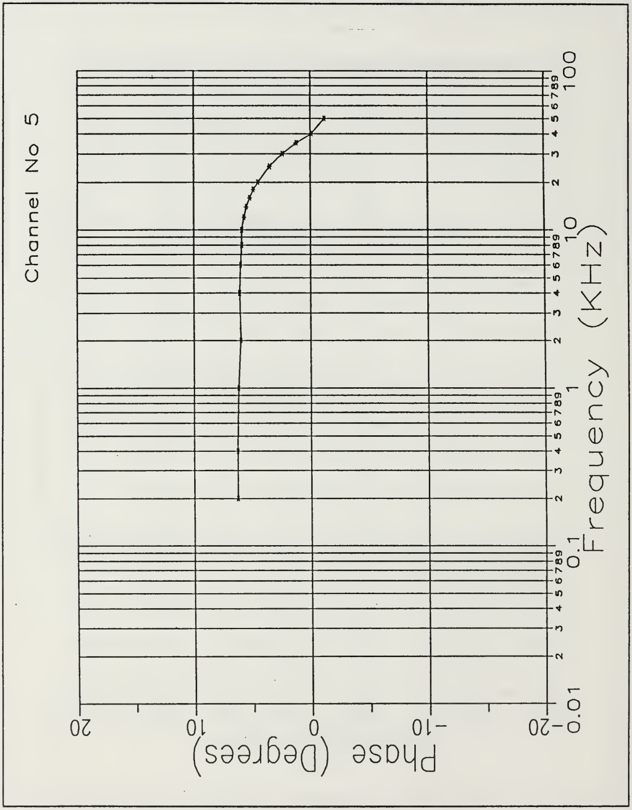

Figure 45. End-To-End Phase Characteristic for Channel 5.

the output optical port. The attenuator was connected in the path between the optical transmitter (HFBR-1404) and the optical receiver (HFBR-2404). The attenuator has an insertion loss of 3.5 dB.

The fiber used was a 100/140 μm multimode graded-index fiber manufactured by Chromatic Technologies Inc. This fiber has a numerical aperture NA > 0.25, a bandwidth-distance product 200 MHz-km and attenuation 3.9 dB/km at 820 nm. The optical transmitter chosen provides −17.5 dBm of power. Having all five channels transmitting and increasing the attenuation with the attenuator, it was found that good quality was maintained for 11 dB attenuation for the weakest channel and 13 dB for the strongest. Therefore, signal reception requires −28.5 dBm of received power. Knowing that the fiber has a loss of 3.9 dB/km and that the insertion loss of the attenuator is 3.5 dB, we can find the range of the system. Thus,

Power output of the transmitter:	−17.5 dBm
Insertion loss of the 1950 attenuator:	3.5 dB
Acceptable attenuation:	11.0 dB
Power for the optical receiver:	−32.0 dBm
Fiber losses:	3.9 dB/Km
Connector losses:	2.0 dB

Therefore the range of the system is

$$R = \frac{14.5 - 2}{3.9} = 3.2 \text{ km} \tag{3.2}$$

Since most of the available fiber cables are in lengths of 1 km, the range of the system is 3 km.

V. CONCLUSION

Analog transmission is still useful for a fiber optic communication link, as proved by the operation of the system. The fact that one can transmit five different channels with five different frequencies in the range from 200 Hz to 4 kHz as voice data or up to 20 kHz (audio signals) with a constant gain is one of the satisfactory results of this thesis. The other impressive result comes from the operation of the active bandpass filter in high frequencies. Therefore the 1-10 MHz band can be useful for the transmission not only five but ten or more audio signals, since there is enough separation between the subcarrier frequencies of the channels. On the other hand the cost, as well as the weight and size of a system like the one designed and tested in this thesis, are greatly reduced since there is no need for passive bandpass filters.

The optical receiver puts a limitation to the number of channels and the spectrum used. The HFBR-2404 analog optical receiver used has a frequency response which is typically DC to 25 MHz. The last number puts an upper limit in frequency bandwidth that can be used, since the maximum number of channels transmitted can be 25 MHz / 40 kHz = 625 channels. The range of the system is also satisfactory but the use of a laser diode as the optical source instead of the LED used could extend it more. This, of course, adds cost to the system so there is a trade-off between range and cost at this point.

Suppressed-carrier systems have the advantage over normal AM in that they require less power to transmit the same information. This is because there is no carrier transmission which can take up to 75 percent of the total transmitted power. The result is less expensive transmitter, but more complex and expensive receiver for AM-SC systems, since there is a need for carrier reinsertion and synchronization during the demodulation process. Application for the present link is envisioned for point-to-point communications where there are only a few receivers for one transmitter and the complexity in the receiver is justified. This

is the reason why the capability of AM-DSB/SC modulation was explored in this project.

Although AM-DSB/SC modulation was used, the successful operation of the filters and the improved quality factor Q permit the use of AM-SSB/SC which is in favor for audio communications. This way half of the bandwidth and less power is needed for the system and it is less sensitive to frequency and phase errors in the local carrier used for demodulation.

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